Controllable mixer

Nowadays, receivers for modulated radio-frequency
signals are normally in the form of heterodyne receivers.
5   Heterodyne receivers use a mixing stage to which the input
signal to be received and an oscillator signal are
supplied. The oscillator signal is tunable as a function of
the desired frequency which is intended to be received. The
mixing stage produces at its output a signal which, for
10 example, is at a lower frequency than the input signal.
This frequency is referred to as the intermediate
frequency. The input signal, after having been down-mixed
to the intermediate frequency, is passed via a bandpass
filter and is demodulated in a downstream demodulator
15 stage, in a typical receiver. The mixer is frequently
preceded by a controllable amplifier which matches the
level of the input signal to the input of the mixer. This
measure prevents interference signals being produced by
overdriving as a result of non-linearities in the mixing
20 stage. On the other hand, weak input signals are amplified
to such an extent that any noise which is added in the
mixer has a negative effect on the signal-to-noise ratio.
The so-called automatic gain control (AGC) thus ensures
that the level of an input signal is matched to a
25 downstream stage.

Furthermore, the signal flowpath upstream of the mixer
also frequently contains a tunable bandpass filter, by
means of which signals which are adjacent to the useful
30 signal are reduced or suppressed. Suppression or reduction
of signals which are adjacent to the useful signal is
necessary because intermodulation interference can be
caused in the mixer by the proximity of the useful signal
and adjacent signals. Furthermore, adjacent signals which
35 are at a higher signal level than the useful signal
overdrive the mixer. This is the situation when the

controllable amplifier upstream of the mixing stage matches
the level of the useful signal to the input of the mixer
and at the same time also raises the adjacent signal above
the maximum permissible input level of the mixer.

5   The suppression of signals which are adjacent to a
useful signal involves a high degree of circuitry
complexity. Bandpass filters upstream of the mixer must be
tunable to the tuned frequency. Furthermore, circuits or
10   filters for suppressing adjacent signals must be trimmed at
the factory during manufacture of receivers.

It is thus desirable to specify a circuit which has a
mixer and in which intermodulation interference is
15   suppressed and the noise behaviour is improved with less
complex circuitry and with less need to trim circuit parts
during manufacture of receivers. A further aim is to
specify a method for controlling a circuit according to the
invention for optimization of the immunity to interference.

20   A mixer such as this and a method such as this are
specified in the independent claims. Advantageous
refinements and further developments are specified in the
dependent claims.

25   The mixer according to the invention has at least one
transistor whose operating point can be set by means of a
control signal. An evaluation circuit which is connected
via a bandpass filter downstream from the mixer evaluates
30   the signal quality of the output signal. If the
intermodulation interference is strong, as occurs, by way
of example, when two strong signals are at closely adjacent
frequencies, the operating point of the mixer is set such
that a high collector current flows. When that collector
35   current is high, the modulation range of the transistor in
the mixer increases. The larger modulation range is

required for two adjacent strong signals in order to avoid
the creation of intermodulation products as a result of
non-linearities in the transistor. One embodiment of the
mixer makes use of an effect which occurs particularly in
bipolar transistors. In this case, the mixing gain of the
transistor at the same time decreases when high collector
currents occur, which are required for large input signals.
When the collector currents are low, the mixing gain rises.
The collector current is reduced for small input signals
and when the level of the adjacent signals is only low,
because the requirements for the modulation range of the
transistor are less. At the same time, a higher mixing gain
occurs in mixers designed using bipolar transistors, and
this is desirable for low input signals. Furthermore, the
noise from the transistor is also reduced when the
collector currents are low.

During reception of digitally coded signals, in which
error correction is possible, the signal quality can be
determined in a simple manner by evaluation of the error
rate. However, other possible ways to determine the signal
quality are also feasible, depending on the type of
modulation used and on the input signal, for example
analysis of the frequency spectrum of the output signal
from the mixing stage.

The invention advantageously allows dynamic matching of
the mixer characteristic to the respective reception
situation, taking into account signals adjacent to the
received signal. This makes it possible to improve the
resistance to interference, with less circuitry complexity.

A receiver circuit according to the invention for
reception of digital signals does not require a tunable
bandpass filter at the input. The digital circuit evaluates

the error rate of the received signal, and controls the characteristic of the mixing stage accordingly.

In particular, the circuit and the method are also suitable for mobile appliances or other appliances in which the current drawn should be minimal (intelligent power management). Dynamic adjustment of the characteristic of the transistor in the mixing stage makes it possible to reduce the collector current, and thus the entire current that is drawn, when the reception situation is good and the signal levels are low. The reduced current drawn advantageously makes it possible to reduce the effort for heat dissipation in the circuitry. This also simplifies integration of the mixer and of further components, such as a demodulator, in a single integrated circuit.

In a further development, initial values for operation of the mixer in the receiver are stored in a memory. These initial values include, for example, information about the modulation method, the code rate and/or the symbol rate. The modulation method may, inter alia, comprise phase modulation methods such as BPSK (Binary Phase Shift Keying), QPSK (Quadrature Phase Shift Keying), 8PSK (8 Phase Shift Keying), or combined phase amplitude modulation methods such as QAM (Quadrature Amplitude Modulation), or else frequency modulation methods such as OFDM (Orthogonal Frequency Division Multiplex).

The initial values are used as the basis for assessment of the current signal quality, and the operating point of the mixer is set so as to achieve at least a desired minimum signal quality. Values for a desired minimum signal quality are advantageously likewise stored in the memory, and the desired minimum signal quality may differ, depending on the modulation method that is used. In this case, values for the desired minimum signal quality are

stored for each modulation method. In the case of digitally coded signals, the signal quality is inversely proportional to the error rate.

5   In a further development, individual optimization routines are stored for each of the different initial values mentioned above. The optimization routines are then used to optimize the setting of the mixer.

10   The theoretical principle of the invention can be derived from analysis of the non-linear transmission characteristic of a four-pole network (illustrated here only up to the 3rd order):

15       $$y = a \cdot x + b \cdot x^2 + c \cdot x^3 \qquad\qquad (1)$$

A two-tone signal x(t) is passed to the four-pole network with the transfer function as in equation (1):

20       $$x(t) = u \cdot \sin(\varpi_1 \cdot t) + v \cdot \sin(\varpi_2 \cdot t) \qquad\qquad (2)$$

Equation (3) is obtained by substitution of (2) and (1):

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$$y = \frac{1}{2} b \cdot u^2 + \frac{1}{2} b \cdot v^2 \qquad\qquad [1]$$

$$+ a \cdot u \cdot \sin(\varpi_1 \cdot t) + a \cdot v \cdot \sin(\varpi_2 \cdot t) + \frac{3}{4} c \cdot u^3 \cdot \sin(\varpi_1 \cdot t) + \frac{3}{4} c \cdot v^3 \cdot \sin(\varpi_2 \cdot t)$$

[2]

$$+ \frac{3}{2} c \cdot u \cdot v^2 \cdot \sin(\varpi_1 \cdot t) + \frac{3}{2} c \cdot u^2 \cdot v \cdot \sin(\varpi_2 \cdot t) \qquad\qquad [3]$$

30       $$+ b \cdot u \cdot v \cdot \cos((\varpi_1 - \varpi_2)t) - b \cdot u \cdot v \cdot \cos((\varpi_1 + \varpi_2)t) \qquad\qquad [4]$$

$$- \frac{1}{2} b \cdot u^2 \cdot \cos(2\varpi_1 \cdot t) - \frac{1}{2} b \cdot v^2 \cdot \cos(2\varpi_2 \cdot t) \qquad\qquad [5]$$

$$-\frac{3}{4}c \cdot u^2 \cdot v \cdot \sin((2\varpi_1 \pm \varpi_2)t) - \frac{3}{4}c \cdot u \cdot v^2 \cdot \sin((\varpi_1 \pm 2\varpi_2)t) \qquad [6]$$

$$-\frac{1}{4}c \cdot u^3 \cdot \sin(3\varpi_1 \cdot t) - \frac{1}{4}c \cdot v^3 \cdot \sin(3\varpi_2 \cdot t) \qquad [7]$$

In the equation above, the equation parts are as follows:

[1] represents the DC component,

[2] represents the linear component,

[3] represents the cross-modulation component,

[4] represents the intermodulation IM2,

[5] represents the square component (twice the frequencies of the signals ω_1 and ω_2);

[6] represents the intermodulation IM3, and

[7] represents the cubic component (three times the frequencies of the signals ω_1 and ω_2).

The factors b and c in (1) can be controlled by suitable control of the mixer characteristic such that the non-linear components IM2 [4] and IM3 [6] are variable. The characteristic, and hence the factors b and c, are controlled via the control input in the circuit according to the invention.

The invention will be described in the following text with reference to the drawing, in which:

Figure 1   shows a receiver according to the prior art,

Figure 2   shows a receiver with a mixer according to the invention,

Figure 3   shows a first schematic illustration of a mixer according to the invention,

Figure 4   shows a second schematic illustration of a mixer according to the invention,

Figure 5   shows an illustration of the intermodulation
            immunity as a function of the collector current
            for a transistor,

Figure 6   shows a schematic illustration of the input and
5   output signals of a mixer for different
            transistor operating points.

        Identical or similar elements are provided with the
same reference symbols in the figures.

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        Figure 1 shows a schematic block diagram of a receiver
according to the prior art. An input signal RF_{in} is passed
to a tunable bandpass filter 1. The tunable bandpass filter
1 is used for selection of the desired input signal, and
15 for suppression of possible adjacent signals. The signal is
passed from the tunable bandpass filter 1 to a variable-
gain amplifier 2. The amplifier 2 is connected to a mixer
3. The mixer 3 is also supplied with the signal at a
variable frequency from an oscillator 4. An intermediate
20 frequency signal IF is produced at the output of the mixer
3 at a frequency which is lower than the frequency of the
input signal RF_{in}. The intermediate frequency signal IF is
passed to a bandpass filter 6 at a fixed mid-frequency. The
intermediate frequency signal is passed from the bandpass
25 filter 6 to a control circuit 7, and to a demodulator 8.
The demodulated signal is produced at an output 9 of the
demodulator 8 for further processing. The control circuit 7
uses a control signal AGC to control the variable amplifier
2. This control loop ensures that the input signal RF_{in} is
30 applied to the mixer 3 at a suitable signal level. The
demodulator 8 demodulates the signal for further
processing.

        Figure 2 shows a schematic block diagram of a receiver
35 with a mixer according to the invention. An input signal
    RF_{in} is passed to an amplifier 2. The input signal is

passed from the amplifier 2 to a mixer 3. The mixer 3 is
supplied with the signal at a variable frequency from an
oscillator 4. Furthermore, the mixer 3 is supplied with a
signal AGQC. An intermediate frequency signal IF is passed
5   from the output of the mixer 3 to a bandpass filter 6 at a
fixed mid-frequency. The signal is passed from the bandpass
filter 6 to a demodulator 8. The demodulator 8 demodulates
the received signal, and produces it at an output 9. The
signal at the output 9 is also passed to an evaluation
10 circuit 7, which assesses the signal quality and produces
the monitoring signal AGQC as a function of the quality of
the received and demodulated signal, which monitoring
signal is applied to the mixer 3. A memory 5 is connected
to the evaluation circuit for storage and for reading
15 initial values, as well as data obtained during operation.

Figure 3 shows a first schematic circuit diagram of a
mixer according to the invention. A radio-frequency signal
RF_{in} is passed via a coupling capacitor 11 to the base
20 connection of a transistor 12. The operating point of the
transistor 12 is set via a voltage divider comprising the
resistors 13 and 14 at the base connection of the
transistor 12. A control voltage U_S is applied to the base
connection of the transistor 12 via a resistor 16. The
25 control voltage U_S is derived from the signal AGQC, which
is not illustrated in the figure. The operating point of
the transistor 12 can be varied by means of the control
voltage U_S. A capacitor 18 and an inductance 19, connected
in parallel, are connected to an operating voltage UB at
30 the collector connection of the transistor 12. The parallel
circuit formed by the capacitor 18 and the inductance 19
forms an IF filter 17. The intermediate frequency signal IF
is also produced at the collector output of the transistor
12, and is emitted via a coupling capacitor 23. An emitter
35 resistor 22 is connected to earth at the emitter connection
of the transistor 12. Furthermore, the emitter connection

of the transistor 12 is supplied via a coupling capacitor
21 with the signal LO at a variable frequency from an
oscillator which is not illustrated in the figure.

Figure 4 shows a second schematic illustration of a
mixer according to the invention. The mixer in Figure 4 is
suitable for processing balanced signals. The negative
mathematical sign in the annotation of the signals
indicates that the signals are in antiphase. A signal RF_{in}
is passed via a coupling capacitor 11 to the base
connections of two transistors 26 and 29. The base
connections of the transistors 26 and 29 are connected to
earth via a resistor 14. An antiphase signal $-RF_{in}$ is
passed via a coupling capacitor 111 to the base connections
of two transistors 27 and 28. The base connections of the
transistors 27 and 28 are connected to earth via a resistor
114. The transistor pairs 26 and 27 as well as 28 and 29
are connected as differential amplifiers. The emitters of
the transistor pairs 26 and 27 as well as 28 and 29 are
respectively connected to one another. The connected
emitter connections of the transistors 26 and 27 are
connected to earth via a resistor 30 and a capacitor 31.
The connected emitter connections of the transistor pair 28
and 29 are likewise connected to the capacitor 31 via a
resistor 130, and are connected to earth via this capacitor
31. The signal LO from an oscillator is applied to the
connected emitter connections of the transistors 26 and 27
via a coupling capacitor 21. The antiphase signal $-LO$ is
applied to the connected emitter connections of the
transistor pair 28 and 29 via a coupling capacitor 121. A
control voltage U_s, which is derived from the signal AGQC
(which is not illustrated in the figure), is connected
between the resistor 30 and the capacitor 31. The operating
points of the differential amplifiers formed by the
transistor pairs 26 and 27 as well as 28 and 29 can be
adjusted by means of the control voltage U_s. The collector

connections of the transistors 26 and 28 are connected to
one another. The collector connections of the transistors
27 and 29 are likewise connected to one another. The
intermediate frequency signal IF and the associated
5   antiphase signal -IF can be tapped off at the connected
collector connections of the transistors via output
capacitors 23 and 123. A parallel circuit formed by an
inductance 19 and a capacitor 18 is coupled between the
connected collector connections of the transistor pairs 26
10   and 28 as well as 27 and 29. The circuit formed by the
inductance and the capacitor forms an intermediate
frequency filter 17. In Figure 4, the inductance 19 is
formed from two series-connected inductance elements, at
whose centre connection the supply voltage for the
15   differential amplifiers is fed. Feeding the supply voltage
via the centre connection avoids the direct current having
any influence on the inductance.

Figure 5 shows an illustration of the intermodulation
20   resistance IM3 of a transistor, as a function of the
collector current. The family of characteristics clearly
shows that the intermodulation immunity level is a function
of the collector current when the collector/emitter voltage
is constant.

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By way of example, Figure 6 shows a simplified
schematic illustration of the input and output signals of a
mixer for different operating points of a transistor.
Figure 6a shows two inputs signals RF_{use} and RF_{adj} with the
30   same signal levels. The useful signal RF_{use} is at a
frequency of 205 MHz. The adjacent signal RF_{adj} is at a
frequency of 214 MHz. It is assumed that the mixer
oscillator is operating at a frequency of 200 MHz. The
intermediate frequency signals IF_{use} and IF_{adj}, on the one
35   hand, and the undesirable intermodulation product IF_{int}, on
the other hand, are produced in the mixer. The useful

intermediate frequency signal IF_{use} is at a frequency of
5 MHz (205 MHz - 200 MHz), the adjacent intermediate
frequency signal is at frequency of 14 MHz (214 MHz -
200 MHz), and the interference intermediate frequency

5   signal IF_{int}, which is formed by intermodulation, is at a
frequency of 4 MHz (200 MHz - (2 × 205 MHz - 214 MHz)).
Figure 6b shows examples of useful, adjacent and
interference intermediate frequency signals. The diagram in
Figure 6b is based on the assumption that the mixing

10 transistor is set to a high mixing gain. The three
intermediate frequency output signals are at relatively
high levels, with the interference intermediate frequency
signal IF_{int} being at only a slightly lower level than the
useful intermediate frequency signal IF_{use}. The

15 intermodulation separation, which is the separation between
the useful signal and the interference signal, is, by way
of example, assumed to be X dBc, where dBc represents a
weighted measurement. Figure 6c shows the output signals
from the mixer when the mixing gain of the mixer transistor

20 is lower. The useful intermediate frequency signal IF_{use} is
at a lower level than in Figure 6b. The interference
intermediate frequency signal IF_{int} has not been reduced to
the same extent as the useful intermediate frequency
signal. The intermodulation separation was increased

25 considerably in comparison to the example in Figure 6b, and
is Y dBc. In this case, Y dBc is greater than X dBc.